

**ACTIVE LOAD ARRANGEMENT****TECHNICAL FIELD OF THE INVENTION**

The present invention relates to an active load arrangement  
5 providing proper output load to a device under test.

**DESCRIPTION OF RELATED ART**

Automatic test equipment is frequently employed to run  
diagnostic tests on integrated circuit devices. A purpose of  
10 a testing device is to ensure that integrated circuits under  
test provide the proper output voltages given the input  
stimulus of the tester. It is necessary that such devices  
simulate the proper output load. In the US patent US  
4,712,058 is disclosed an active load network for a device  
15 under test and for turning on either a current source or a  
current sink to properly load its output. The current sink  
and current source each comprises a pair of CMOS transistors  
connected in series. An active load network that provides  
low resistance for direct current DC and high impedance for  
20 alternating current AC is often necessary when testing an  
object of for example line interface type. This type of load  
network prevents superposed AC parameters to be essentially  
influenced by DC load. When measuring AC parameters of a  
test object, the AC impedance of the load network must be  
25 essentially higher than the AC impedance of the test object.  
High AC impedance and low DC resistance can be accomplished  
by an inductance. However, when measuring on low frequency  
signals from about 200 Hz and high impedance load above 150  
kohm is required, the inductance has to be more than 120 H.  
30 This type of AC impedance is necessary when measuring for  
example on line interface circuits within POTS (Plain Old  
Telephone Service) telephony. A passive inductance i.e. a

coil of that inductance size which can stand enough current without being saturated will be bulky and expensive. The necessary size of such a coil may cause other types of problems like magnetic energy storing causing high voltages at circuit breaking or other rapidly changing loads. Also magnetic and electrical EMC problems depending on internal and external crosstalk may occur. Another possible solution would be to design an active inductance. Active inductances have been made with bipolar transistors. Since bipolar transistors are current controlled devices, several are needed in some kind of high current gain configuration like Darlington connection to get the necessary high impedance to the control input of the circuit. Solutions like the one above is however complicated.

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#### **SUMMARY OF THE INVENTION**

The present invention relates to a problem how to provide proper output load to a device under test requiring low resistance for DC voltages and high impedance for AC voltages.

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A purpose with the invention is to facilitate measurements of a test object without essentially influencing superposed AC voltage parameters from the device under test.

The problem is solved by the invention by an active load arrangement comprising a voltage-controlled transistor associated with a feedback arrangement. The feedback arrangement controls the DC bias point, and by superimposed AC input the arrangement controls the phase and the amplitude for the control signals to the transistor inputs.

More in detail, the active load arrangement comprises a voltage-controlled transistor called Metal Oxide Semiconductor Field Effect Transistor, MOSFET, having a

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source, a gate and a drain. The drain is connected to an input of the arrangement associated with an output of the object under test. A resistor is connected between the gate and the drain. An inductance is connected between the source and an output of the arrangement associated with an input of the object under test. A capacitance is connected between the gate and the arrangement output.

One advantage with the invention is that the impedance for the active load arrangement is not limited to the internal drain-source resistance but can be further enhanced by the feedback nets.

Another advantage with the invention is that high impedance is obtained without the need of bulky passive components.

Yet another advantage is that the MOSFET is a cheap component and only one active device with a few additional components are needed for the invention.

The invention will now be described more in detail with the aid of preferred embodiments in connection with the enclosed drawings.

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#### **BRIEF DESCRIPTION OF THE DRAWINGS**

Figure 1 shows a schematic diagram of an active load arrangement connected to a test object.

Figure 2 shows an active load arrangement more in detail with the aid of an equivalent MOSFET diagram.

Figure 3 shows results from a simulation using the active load arrangement.

Figure 4 shows a schematic diagram of two active load arrangements connected in reverse.

**DETAILED DESCRIPTION OF EMBODIMENTS**

In figure 1 is disclosed a test arrangement TA comprising an active load arrangement Z connected to a test object TO. The active load arrangement provides proper output load to the object under test. The test object, which for example can be a POTS line interface circuit in a telecommunication system, comprises an object output O1 and an object input I1. The load arrangement Z comprises an arrangement input I2 associated with the object output O1. The arrangement Z also comprises an arrangement output O2 associated with the object input I1. The load arrangement Z further comprises a MOSFET with a drain D, a gate G and a source S. The semiconductor in this example is of the type IRF610. The drain D is connected to the arrangement input I2. A first resistance R1, in the example 2,2E6 ohm, is connected between the drain D and the gate G. A second resistance R2, in the example 2,2E6 ohm, is connected between the gate G and the source S. A capacitor C1, in the example 6,8  $\mu$ F, is connected between the gate and the arrangement output O2. An inductance L1, in the example 42mH, is connected between the source and the arrangement output O2. The inductance L1 is called a first feedback net FBN1 and the resistances R1 and R2 together with the capacitor C1 is called a second feedback net FBN2.

The test object TO in this example is a POTS line interface circuit and it works under normal OFF HOOK operation with an AC voltage superposed on a DC voltage. In order to activate and measure the AC parameters of the test object, a load arrangement called DC-loop is necessary to supply low-resistance DC load and high-impedance AC load. The low-resistance DC load make the line interface circuit to go into the state where it is working with 2 wire AC voice signals. The high-impedance AC-load is needed because the superposed AC parameters must be as uninfluenced as possible

by the DC load. The load arrangement must be essentially higher than the AC impedance of the test object TO.

The invention will now be explained with the aid of figure 1 and figure 2. Figure 2 discloses more in detail the active load arrangement Z shown in figure 1. Figure 2 shows the earlier mentioned first resistance R1, the second resistance R2 and the capacitor C1. In figure 2, also the internal resistance RL of the inductance L1 is shown. The MOSFET shown in figure 1 is shown in figure 2 as an equivalent circuit diagram suitable for simulation of the invention, i.e. a simulation model. A voltage controlled Current Source CS is situated in the MOSFET model between the gate G, the drain D and the source S. A static drain-to-source resistance  $R_{ds}$  is shown between drain and source. A reverse transfer capacitance  $C_{rss}$  is shown between the gate and the drain. To be able to simulate the approximate DC bias point of the circuit a voltage bias  $V_{bias}$  is placed between the source S and a minus pin of the voltage controlled current source CS.

A suitable working-point for the MOSFET is selected with respect to the DC needed for the test object TO, the reverse transfer capacitance  $C_{rss}$  and the power dissipation of the MOSFET. High drain D to source S voltage results in high power dissipation and low  $C_{rss}$ . Low drain current results in high  $R_{ds}$ , low forward transconductance  $G_{fs}$  and low power dissipation. The working point is selected by means of the first resistance R1 and the second resistance R2. R1 and R2 will bias the gate so that the drain voltage will be approximately 10V and so that the gate voltage will be approximately 5V when using a MOSFET type IRF610 which has a  $V_{bias}$  of about 5V for the current values that are valid, i.e. 10-100mA. This bias will create sufficient drain-source voltage for different currents to the testobject. The

capacitor C1 together with the resistance R1 and R2 will have the effect of a low-pass filter. Since the MOSFET is a voltage controlled device it has a very high input resistance. R1 and R2 can then be set to very high  
 5 resistance values without being loaded by gate currents. This is necessary because R1 and R2 add AC load to the circuit. C1 can be set to a low capacitance value while the circuit still will reach the necessary filtering capability, because R1 and R2 are high resistive. This is also an  
 10 advantage because the low capacitance value eliminates the need of an electrolytic capacitor.

The superposed AC delivered from the test object will due to the low-pass filter R1, R2 and C2 cause the AC voltage on the gate G to go towards the object output O2.

15  $R_{ds}$ , CS and L1 creates a signal path to the source S. The superposed AC will then result in a voltage across L1 which gives a source to gate voltage when the voltage between the gate and the output O2 goes towards zero which will rise the  $Z_{ds}$  above the  $R_{ds}$  value.

20 The AC behavioral of the MOSFET, somewhat simplified without the affection of  $C_{rss}$ , the impedance  $Z_{ds}$  between drain D and source S can be expressed in the following equation:

$$Z_{ds} = \frac{V_{ds}}{\frac{V_{ds}}{R_{ds}} + Gf_s * V_{gs}}$$

$V_{ds}$  is the AC voltage between drain and source.

25  $R_{ds}$  is the linearized resistance between drain and source at a fixed DC current and voltage. IRF610 is typical 10kohm at DC drain current 80mA and DC drain to source voltage 10V.

$V_{gs}$  is the AC voltage between gate and source.

$G_{fs}$  is the linearized forward transconductance at a fixed DC current and voltage. IRF610 is typical 0.31 A/V at DC drain current 80mA and DC drain-source voltage 10V, for  $V_{gs}=V_{ds}$  then  $Z_{ds}=1/G_{fs}$ . For IRF610 10V/80mA,  $Z_{ds}=3.2$  ohm.

- 5 When the  $V_{gs}$  voltage goes towards zero it can be seen in the equation that drain-source impedance  $Z_{ds}$  goes towards  $R_{ds}$ , for IRF610 10V/80mA,  $R_{ds}=10\text{kohm}$ .

If then  $V_{gs}$  is inverted and  $V_{ds}/R_{ds}+G_{fs}*V_{gs}$  becomes zero,  $Z_{ds}$  is equal to infinity.

- 10 In the invention the feedback nets are arranged so that at low frequencies  $V_{gs}=V_{ds}*R_1/(R_1+R_2)$  applies. The low frequency AC impedance will then be low. At high frequencies the  $V_{gs}$  will change phase and amplitude in a way that the impedance rises and  $V_{ds}/R_{ds}+G_{fs}*V_{gs}$  goes towards 0. When  $V_{gs}$  gets  
15 negative the impedance will be even higher than  $R_{ds}$ .

According to the invention, when the AC gate voltage goes towards zero, the AC drain current  $I_{drain}$  will decrease. Due to the  $R_{ds}$  and  $G_{fs}$ , a voltage  $U_L$  will arise over  $L_1$  and  $R_L$  and an inverted gate-source voltage appears. As can be seen  
20 in the equation, the inverted gate-source voltage  $V_{gs}$  leads to increased drain-source impedance  $Z_{ds}$ , above  $R_{ds}$ . This appears distinctly in figure 3 where a simulation using the exemplified arrangement is presented.

- In figure 3 is disclosed a simulation of the above described  
25 active load arrangement. It is to be observed that the simulation is disclosed schematic with a purpose to show the principle and enhance the understanding of the invention. In figure 3 can be seen three diagrams. A first diagram  $U$  shows

a voltage  $U_G$  between the gate G and the output O2, and a voltage  $U_S$  between the source S and the output O2, as a function of the frequency  $f$ . In a second diagram  $P_h$  the phase between the input I2 and the output O2 is shown as a function of the frequency  $f$ . A third diagram  $Z$  shows the I2 to O2 impedance as a function of the frequency  $f$ . In the first diagram  $U$  can be seen how the source voltage  $U_S$  and the gate voltage  $U_G$  decreases by increasing frequency  $f$  due to the two feedback nets FBN1 and FBN2. At a specific point  $U_S = U_G$  applies. Because of phase differences of about fifty degrees between FBN1 and FBN2,  $U_{SG}$  is almost equal to  $U_S$  or  $U_G$ . At this point the I2 to O2 impedance  $Z$  is higher than  $R_{ds}$ . This can be seen in the third diagram where  $Z = 18\text{kohm}$ . At this point the drain to source impedance  $Z$  is equal to  $R_{ds}$ , as have been explained above. In the second diagram  $Ph$  the phase goes from approximately  $-100$  degrees to  $+100$  degrees by increasing frequency, i.e. from inductive to capacitive impedance. When the phase curve reach zero degrees, the impedance reaches it maximum at the parallel resonance of the active load arrangement. The  $Z$  in the third diagram is resistive at DC and gradually goes inductive from low frequencies up to a point where it creates a parallel resonance circuit together with  $C_{rss}$ . R1 is in parallel with the circuit and limits the maximum impedance that can be reached because at high frequencies the reactance of C1 is much lower than the resistance of R1 and R2.

When the invention is to be used for measurement purposes it may be made for bipolar operation. The invention can be placed in a rectifier bridge on the rectified output side. The rectifier bridge add 2 reversed biased diodes with 2 diode capacitances and the other 2 forward biased diodes add



2 diode voltage drops and diode resistances to the invention.

A second embodiment is disclosed in figure 4. The figure shows a solution where two active load arrangements like the one shown in the first embodiment are connected in reverse and thus bypassed by an internal drain-source diode of one of the MOSFETs. By connecting the two arrangements reversed together and by slightly modify the solution they will be able to share the same inductance L1. This modification is disclosed in figure 4. A load arrangement Z1 comprises a second semiconductor MOSFET2 comprising a second source S2, a second gate G2 and a second drain D2. The second source S2 is connected via a fourth resistance R4 to the second gate G2. The second gate G2 is connected via a fifth resistance R5 to the second drain D2 and to an arrangement output O3. The second drain S2 is connected via the capacitance C1 discussed in the first embodiment, to the gate G. The second gate G2 is connected via a second capacitor to the source S of the earlier discussed MOSFET. The second source S2 is connected to the earlier mentioned impedance L1. Compared to the bipolar rectifier solution this gives no additional load capacitance and only one additional diode voltage drop. However this solution needs more DC bias point stabilisation time than the previous by changing polarity of the input.

Different variations are of course possible within the scope of the invention. The resistor R2 can for example be omitted if lower impedance depending on higher  $C_{rss}$  and a smaller signal before clipping can be accepted. R2 may be connected across C1 instead of between MOSFET gate-source. That will only slightly affect the DC bias voltage depending on the size of RL, which normally is made low to get low different DC voltages across the L1 inductor at different drain currents. The more the DC voltage to source-O2 changes, C1 needs more reloading time, which is not desirable. The

invention is not limited to the above described and in the drawings shown embodiments but can be modified within the scope of the enclosed claims.